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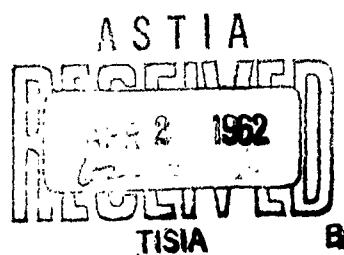
NRL Report 5740

AN L-BAND DIGITAL PHASE SHIFTER

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Search Radar Branch
Radar Division

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U. S. NAVAL RESEARCH LABORATORY
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ABSTRACT

A digitally controlled phase shifter has been developed which has potential application to fixed-array radar systems. The device is capable of producing shifts in phase in steps of 2-13/16 degrees from 0 to 360 degrees. The phase shifter, which operates throughout at 1200 Mc, utilizes G.E. type 6299 uhf-shf triodes, and different lengths of strip transmission lines, in seven stages to produce 128 steps of phase shift. A phase-shift accuracy of ± 3 degrees has been maintained at the center frequency with a power gain of 14 to 17 db. A maximum phase error of ± 20 degrees has been observed over a 10-Mc bandwidth centered at 1200 Mc. Within certain rather narrow limits of frequency, triode grid and plate voltages, rf drive power, and tube characteristics, the circuit is capable of giving excellent control of phase shift. Once the phase shifter is tuned, there is no measurable change in phase with time.

PROBLEM STATUS

This is a final report on one phase of the problem. Work on other phases is continuing.

AUTHORIZATION

NRL Problem R02-33
BuShips Project SF 001-02-02, Task 6085

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AN L-BAND DIGITAL PHASE SHIFTER

INTRODUCTION

The present trend in electronically scanned fixed-array radar systems is toward the use of a large number of radiating elements, each fed by a comparatively low-power transmitter. The need thus arises for a large number of dependable, easily reproduced phase shifters. The use of many components in a fixed-array system dictates that all components be compact and insensitive to mechanical vibration and changes in temperature and power level. Phase shifters for such systems must be simple in operation and control, and relatively inexpensive.

The phase shifter described in this report employs several triode stages to introduce different lengths of transmission line. Each triode stage consists of four sections: power divider, phase-determining transmission lines, triode switch for each of two path lengths, and power-combining plate resonant circuits. The triode stages are cascaded, with each successive stage having different phase-determining lines. The device may be used in receiving arrays or in the low-level portion of transmitting arrays.

PHASE-SHIFT TECHNIQUE

The digital switching circuit of Fig. 1 has definite advantages in simplicity of construction and programming. Actual line lengths are switched in and out to produce the desired relative phase change between input and output. With this arrangement any phase delay from 0 to 360 degrees can be obtained in 128 steps of $360/2^7$, or $2-13/16$ degrees. A G.E. type GL-6299 uhf-shf triode was chosen for the switch, and sufficient gain was realized to overcome the insertion loss of the transmission lines.

Resonant circuits for the 6299 triodes, connecting lines, and power dividers were fabricated as strip transmission line (1,2). The result is a compact, inexpensive phase shifter, easily reproduced using printed-circuit techniques.

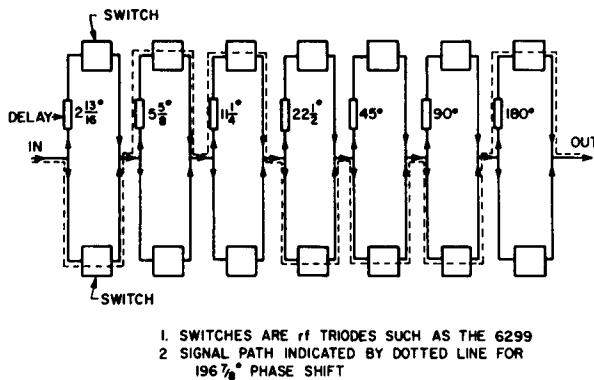


Fig. 1 - Digital phase shifter block diagram

CIRCUIT

The 6299 amplifiers used in the digital phase shifter are connected in a grounded-grid configuration. A center frequency of 1200 Mc was chosen for the operating band. The cathode circuit of a single tube is represented in Fig. 2. An open-circuited half-wavelength resonant line is used, shortened at one end by the grid-to-cathode interelectrode capacity of the tube and at the opposite end by a tuning capacitance, C_T , according to the relations (3):

$$l_{gk} = \frac{\lambda}{2\pi} \tan^{-1} (\omega C_{gk} Z_0)$$

$$l_T = \frac{\lambda}{2\pi} \tan^{-1} (\omega C_T Z_0).$$

A quarter-wavelength section of 70.7-ohm line provides a means by which an external cutoff pulse may be applied to the cathode of the tube. A 0.001-microfarad capacitor located a quarter-wavelength from the resonant line presents a high impedance at the cathode line. The 70.7-ohm line taps the cathode line at a voltage minimum, further reducing leakage of the 1200-Mc center frequency into the cutoff voltage supply. The tap point of the 50-ohm input line was determined experimentally by noting input VSWR and gain as the tap point was moved along the cathode line. The relatively low input impedance of the grounded-grid triode loads the half-wavelength resonant-line section such that a bandwidth of 100 Mc was measured for the cathode circuit.

Figure 3 represents the plate circuit of a pair of 6299 amplifiers. A half-wavelength line is used, shortened by the plate-to-grid interelectrode capacity of an "off" tube at one end, the capacity of an "on" tube at the other end, and two tuning capacitors, C_T , near both ends. Plate voltage is applied at the center of the resonant line near the voltage minimum of the line. The proper choice of this tap point and the choking action of a 5-kilohm,

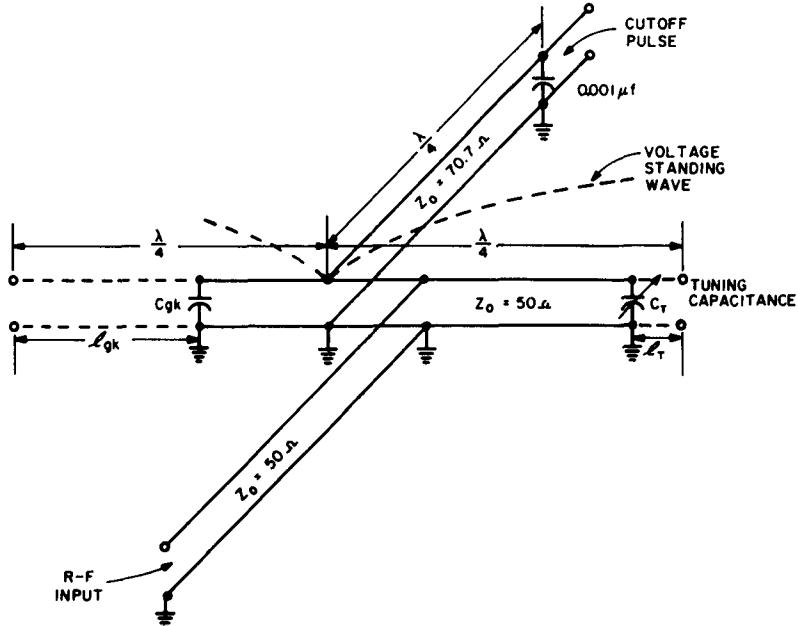


Fig. 2 - Cathode circuit of 6299 amplifier

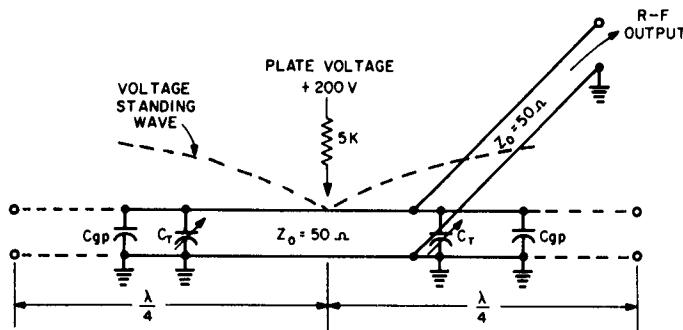


Fig. 3 - Plate circuit of 6299 amplifier

1/2-watt resistor prevent rf leakage into the plate supply. The tap point for the 50-ohm output line was moved along the plate line until a bandwidth of 80 Mc was obtained. A power gain of 5 to 6 db was realized at this point.

Inherent differences in the characteristics of 6299 tube types were compensated for by the tuning capacitors at the end of the cathode line and along the plate line. One or both of the plate tuning capacitors may be used, with the result that some control of bandwidth and gain is possible.

The strip-transmission-line parameters of Fig. 4 were used in the feed lines and resonant circuits of the 6299 amplifiers. Teflon with a dielectric constant, ϵ_r , of 2.1 was used as the dielectric medium for its low-loss characteristics. Ground plates are 1/32-in. silver-plated brass. The center conductors were etched in two halves as mirror images and oriented face-to-face with a total thickness, t , of 2.7 mils. Each half of a center conductor was etched from one-ounce copper-clad fiberglass of 5-mil thickness (4). The ground-plate spacing, b , was determined by the dimensions of the 6299 tube. The width, w , of the center conductor for a desired characteristic impedance was calculated from (5):

$$w = (b - t) \left(\frac{94.15}{\sqrt{\epsilon_r} Z_0} - \frac{C'_f}{0.0885 \epsilon_r} \right),$$

where C'_f is the fringe field capacitance, determined by:

$$\frac{C'_f}{0.0885 \epsilon_r} = \left\{ \frac{2}{(1 - t/b)} \ln \left[\frac{1}{(1 - t/b)} + 1 \right] - \left[\frac{1}{(1 - t/b)} - 1 \right] \ln \left[\frac{1}{(1 - t/b)^2} - 1 \right] \right\} \mu\mu\text{f/cm.}$$

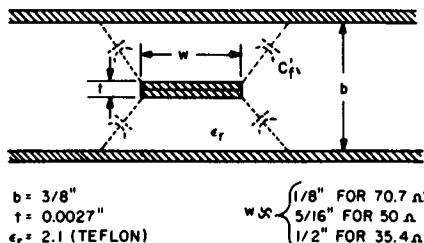


Fig. 4 - Strip transmission line parameters

$$b = 3/8" \quad t = 0.0027" \quad \epsilon_r = 2.1 (\text{TEFLON})$$

$$w \propto \begin{cases} 1/8" \text{ FOR } 70.7 \Omega \\ 5/16" \text{ FOR } 50 \Omega \\ 1/2" \text{ FOR } 35.4 \Omega \end{cases}$$

The cathode tap point of the "off" tube presents a somewhat higher than 50-ohm impedance to the 50-ohm line; the magnitude of this impedance is dependent upon the level of cathode-to-grid bias applied. It was necessary to arrange the power divider so that reflections from this tap point would have a negligible effect on the phase and gain of the "on" tube. Favorable characteristics were obtained by biasing the "off" tube so that the input impedance of the cathode circuit is in the range of 300 ohms, and by using the strip-transmission-line configuration of Fig. 5. One port and a wavelength section of a hybrid ring were replaced by a 100-ohm Filmohm resistor (6,7). This modified hybrid is preceded by a quarter-wavelength transformer of 35.4 ohms characteristic impedance. The measured characteristics of this power divider are plotted in Figs. 6 and 7. The phase variation measured between the input to the quarter-wavelength transformer and the output to the "on" tube has been held to a maximum of 4 degrees over all possible line lengths to the "off" tube. It will be noted that the minimum input VSWR occurs when the line length to the "off" tube is $\lambda/8$, $5\lambda/8$, $9\lambda/8$, etc. This has been chosen as an operating point from which one-half of the desired phase difference, $\theta/2$, has been added for one path and subtracted for the other. Isolation between the branch arms of the power divider has been measured to be greater than 15 db for all line lengths to the "off" tube. The 3-db insertion loss of the power divider reduces the overall gain per stage to 2 to 3 db.

A diagram of one stage is shown in Fig. 8. The cathode and heater lines are overlapping center strips separated by a 5-mil thickness of Teflon tape. A 10-volt, 24-microsecond cutoff pulse is applied to the cathode of the "off" tube, whereas the cathode of the "on" tube is self-biased through a 10-ohm resistor. Switching is accomplished by reversing the connections of the cutoff pulse and 10-ohm resistor. There is an intrinsic phase difference of 180 degrees on the plate line at the rf output as fed by one tube or the other. Feed line L_2 has, therefore, been constructed a half-wavelength longer than L_1 . Hence:

$$L_1 = \lambda(5/8 + \theta/4\pi)$$

$$L_2 = \lambda(9/8 - \theta/4\pi)$$

$$(L_1 - L_2) = \theta\lambda/2\pi - \lambda/2$$

$$\theta = \frac{2\pi}{\lambda} (L_1 - L_2) + \pi \text{ radians.}$$

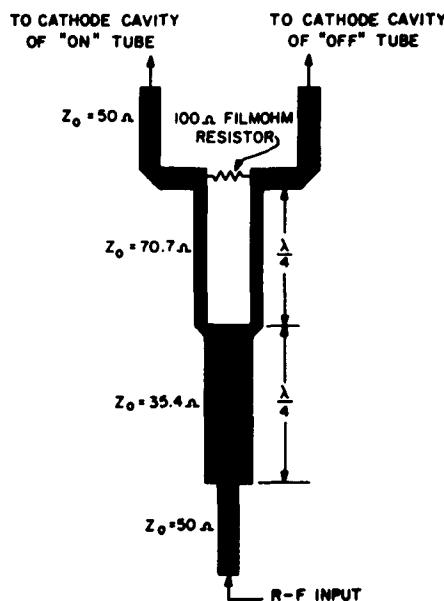


Fig. 5 - Strip transmission line power divider

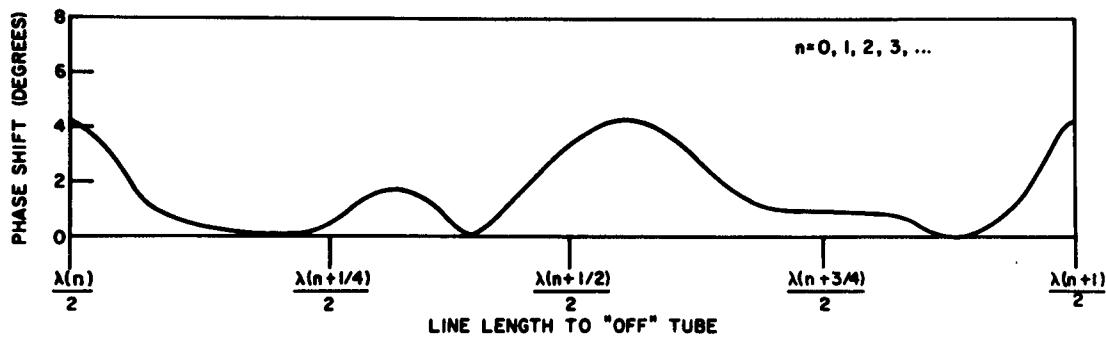


Fig. 6 - Phase characteristics of power divider

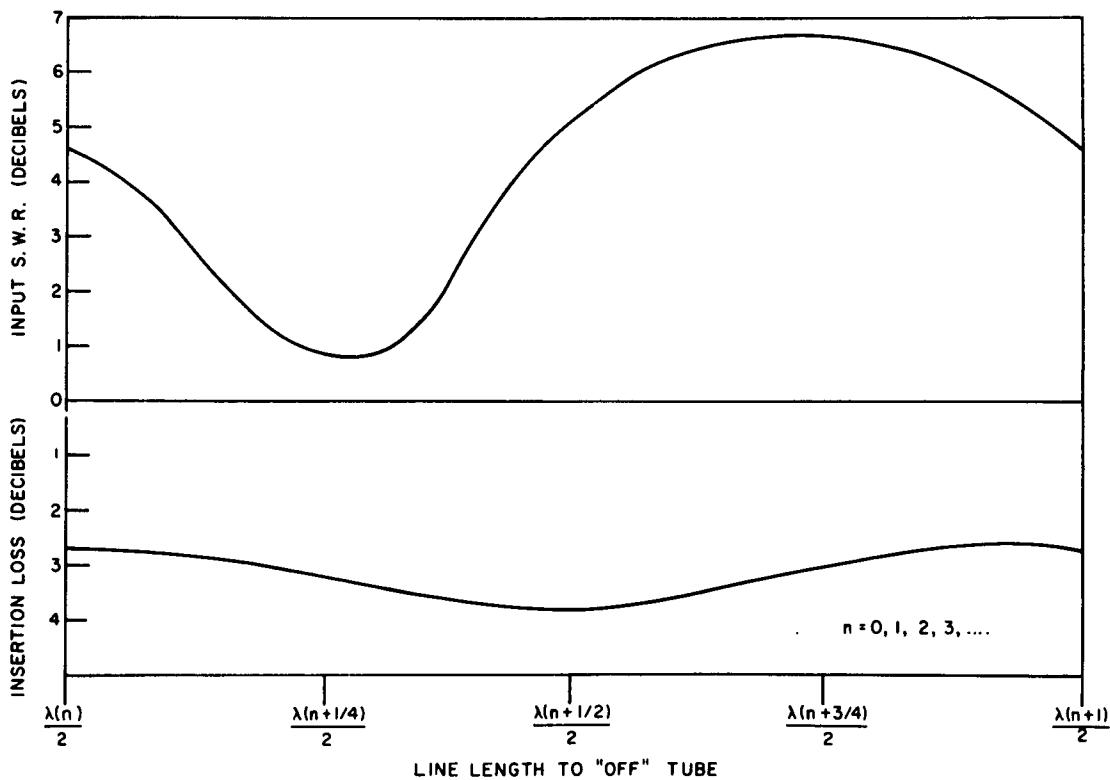


Fig. 7 - Input VSWR and insertion loss of power divider

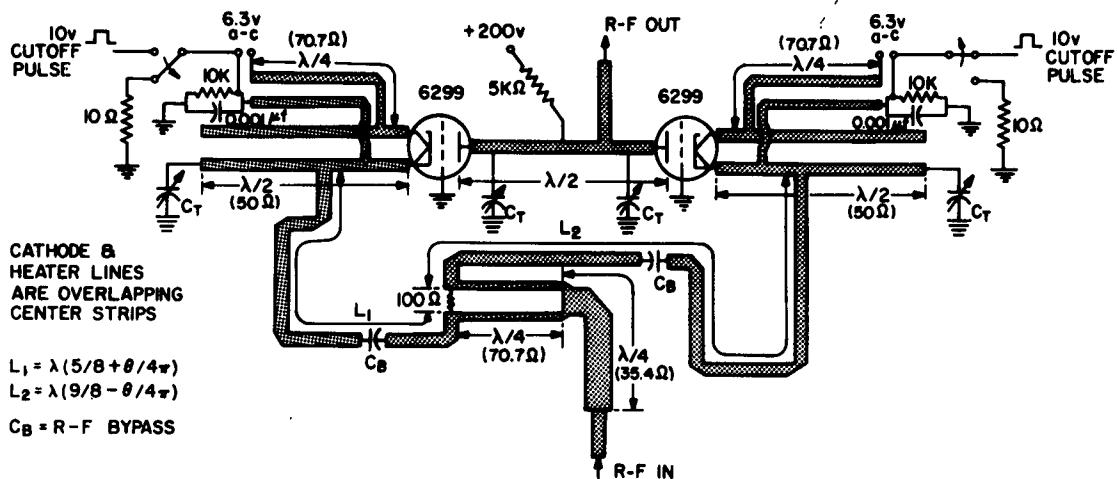


Fig. 8 - 6299 amplifier stage

The rf bypass capacitors, C_B , are formed by etching a gap in one-half of the center conductor and inserting Teflon tape between the top and bottom halves of the center strip for a section of line. Thus, the cathode voltages are isolated from each other and from the plate voltage of the previous stage.

A 45-degree stage is illustrated in Fig. 9. The top ground plate and grid-grounding sections have been removed. Cathode and heater voltages are applied through two-conductor BNC type connectors. The four large screws shown are the tuning capacitors of Figs. 2, 3, and 8. Input and output connections are made by overlapping center conductors and ground plates with those of the adjoining stages. The entire digital phase shifter, composed of the seven stages, is shown in Fig. 10.



Fig. 9 - 45-degree stage

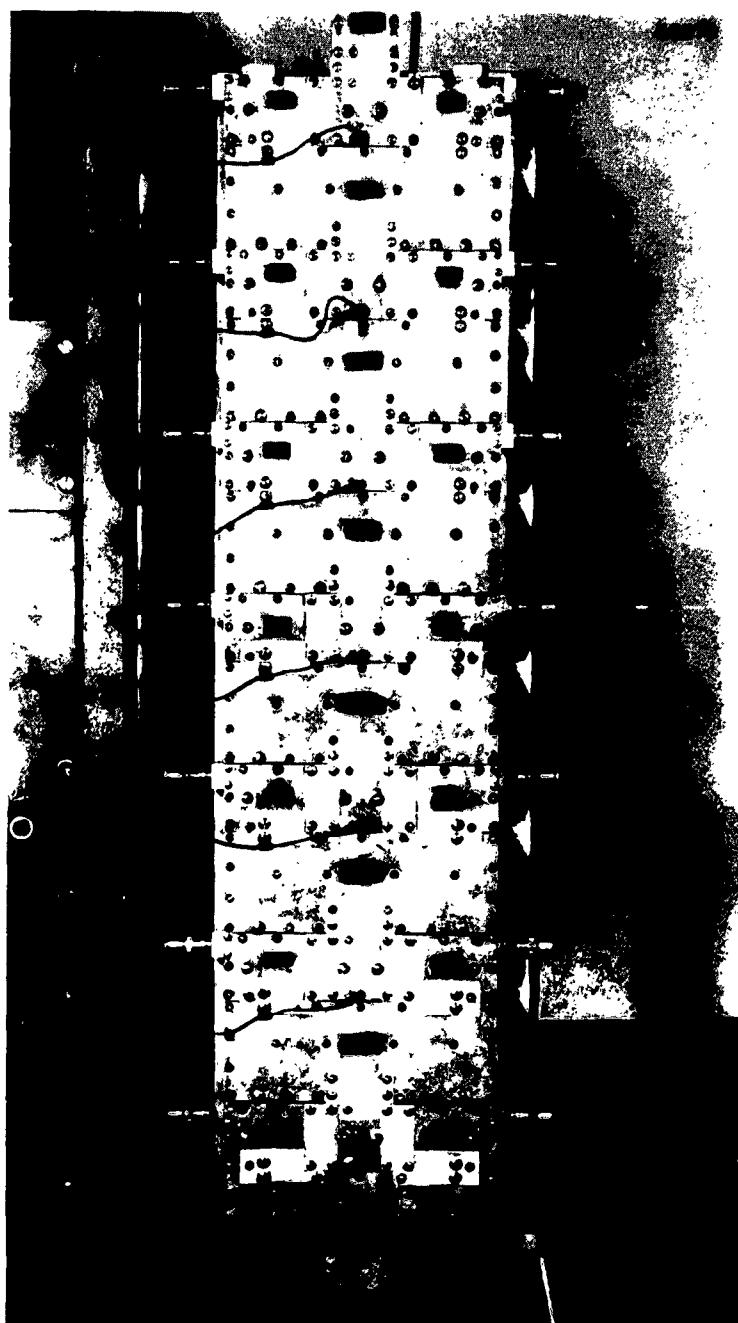


Fig. 10 - Digital phase shifter composed
of seven stages

A 10-microsecond, 1000-pps plate-voltage pulse to a cw-driven preamp provides the rf input to the digital phase shifter. The construction of the preamp is similar to the stage shown in Figs. 8 and 9. Both 6299 tubes are "on" and operate as a conventional push-pull amplifier.

Auxiliary equipment not heretofore mentioned includes a distribution chassis, a pulse modulator, and a modulation trigger source. The pulse modulator provides a 200-volt, 24-microsecond plate pulse as well as a 10-volt, 24-microsecond cutoff pulse. The distribution chassis contains seven manual switches for testing various combinations of the seven stages as well as a plate-current meter and cathode voltage jacks. The cathode current represented by the voltage across the 10-ohm resistor of an "on" tube is illustrated in Fig. 11. The 10-microsecond rf pulse is shown superimposed upon the 24-microsecond, 10-millampere cathode current pulse.



Fig. 11 - Cathode-current pulse

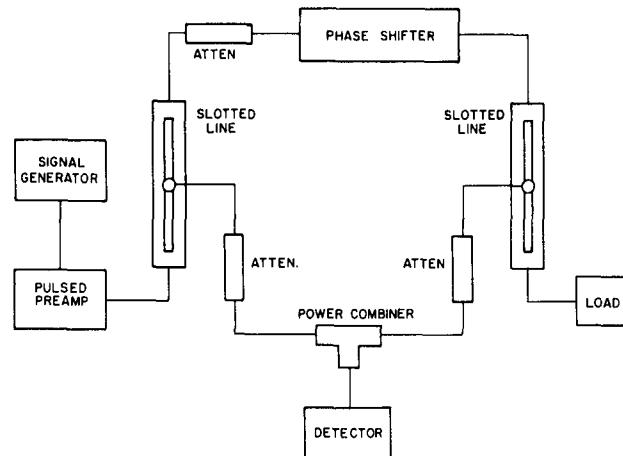
PHASE-MEASURING SYSTEM

Figure 12 is a diagram of the phase-measuring circuit of this experiment. The crystal detectors of the slotted lines have been replaced by shorting slugs, and a crystal detector is placed in the common arm of the power combiner. Attenuators between the probes and power combiner subdue interference from standing waves on these lines.

Signal levels from the probes of the slotted lines to the arms of the power combiner are balanced. One or both of the probes are then moved until a null is indicated in the detector. At this point the signals entering the power combiner are 180 degrees out of phase. A phase change is made in the digital phase shifter. The distance that one of the probes must be moved again to produce a null in the detector represents the number of degrees shift in the digital phase shifter.

All phase-shifting, phase-measuring, and auxiliary equipments are shown in Fig. 13. The slotted lines are Hewlett-Packard Model 805A. A Hewlett-Packard Model 430C microwave power meter is used as a load. The preamp is mounted beneath the seven stages on the side of the equipment rack. At the top of the equipment rack is the detector, a Hazeltine Model 1052 tuned amplifier with a crystal input. Below the detector is the signal generator, a General Radio Type 1218-A unit oscillator. Under the signal generator are the distribution chassis, pulse modulator, and a Hewlett-Packard Model 212A pulse generator which triggers the pulse modulator.

Fig. 12 - Phase-measuring circuit



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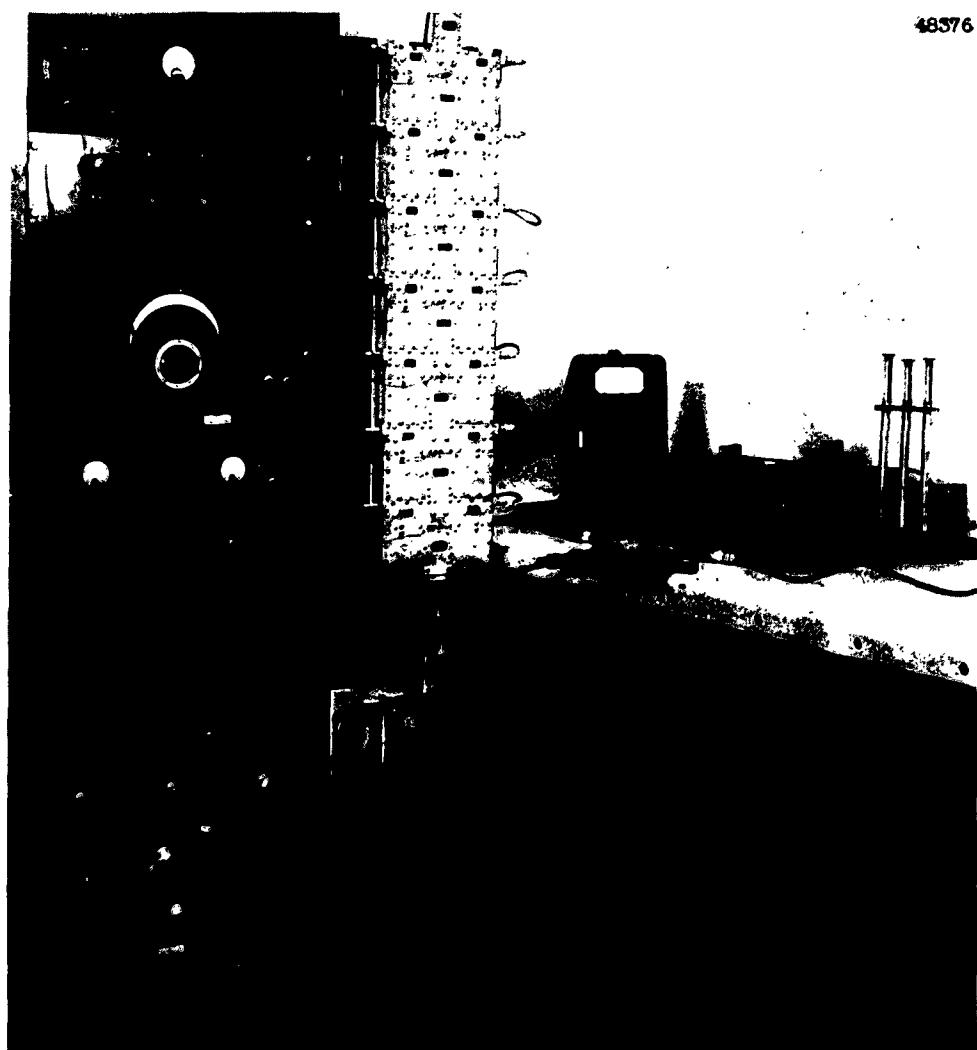


Fig. 13 - Digital phase shifter and auxiliary equipment

MEASURED PERFORMANCE OF DIGITAL PHASE SHIFTER

Phase measurements were made to determine the limitations of the digital phase shifter with variations in rf input power, frequency, plate and cutoff voltages, and timing of the rf pulse relative to the plate-current pulse. The results are described in this section.

The overall bandwidth of the seven 6299 amplifier stages was 20 Mc; therefore the 3-db-gain pass band is from 1190 to 1210 Mc. Figure 14 is a plot of the measured phase-shift error at 1200 Mc. A ± 3 -degree accuracy has been maintained for all individual stages and combinations thereof. A maximum variation of ± 4 degrees was measured as the rf input power was varied from 10 milliwatts peak to 200 milliwatts peak. Lower power inputs result in levels too low for accurate measurement. Higher inputs tend to overdrive the latter stages of the digital phase shifter, resulting in unpredictable phase errors of large order. Excellent accuracy in phase is possible with gains of 14 to 17 db at the center frequency.

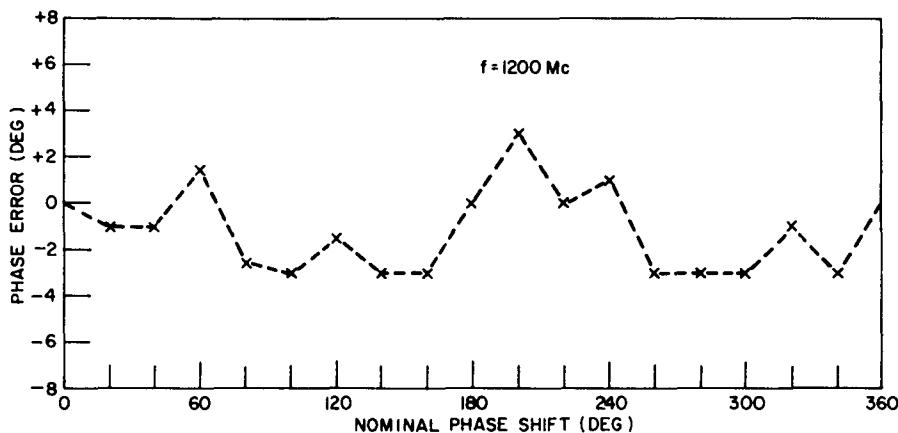


Fig. 14 - Phase measurements at the center frequency

As frequency is varied, a lesser degree of accuracy is possible. The measured phase at 5 Mc to either side of 1200 Mc has been plotted in Fig. 15. With the exception of the 90-degree stage, the individual stages are accurate within ± 5 degrees. The 90-degree stage has a ± 10 -degree error over the 10-Mc bandwidth. The maximum error observed with combinations of the seven stages is ± 20 degrees. Over the entire 3-db, 20-Mc bandwidth, the individual stages show an accuracy of 15 degrees.

A measurement of phase change as a function of plate voltages indicates that there is sufficient variation in tube parameters to prevent the change in phase due to changing plate voltage from being orderly. A maximum of 2 degrees per 100 volts for the basic phase shifters and 17 degrees per 100 volts for any combination of basic phase shifters was measured.

A cutoff value of 6 to 12 volts was used in the design of the modified-hybrid power divider. Lower values of cutoff voltage are insufficient to hold a tube "off." Higher values of cutoff voltage change the impedance presented by the cathode cavity to the feed line such that the characteristics of Figs. 6 and 7 no longer apply. Low or high values of cutoff voltage produce phase errors of magnitudes too large to tolerate, both in individual states and combinations thereof.

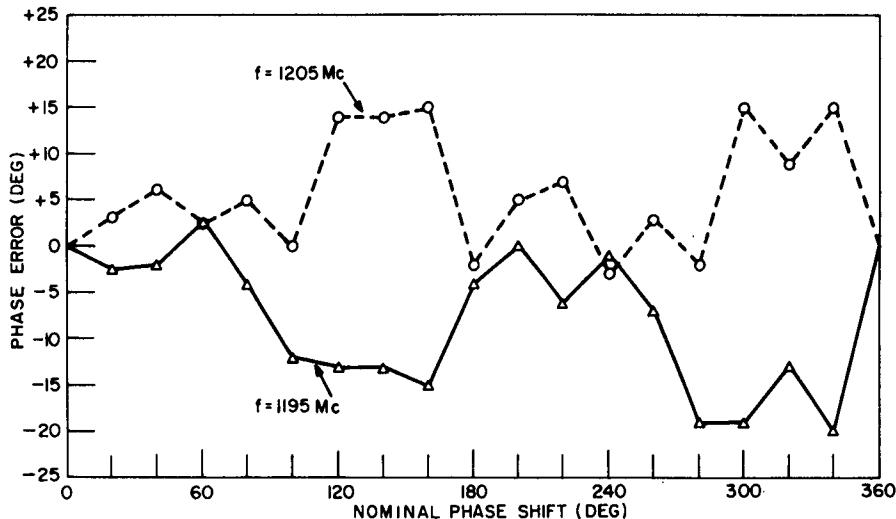


Fig. 15 - Phase measurements over a 10-Mc bandwidth

The 10-microsecond rf pulse is timed to start 5 to 10 microseconds after the beginning of the 24-microsecond plate and cutoff pulses. Thus, the "on" tubes are operating at a predetermined level when the rf pulse is impressed. Actually prepulsing of only 1 or 2 microseconds is necessary, depending on the shape of the plate and cutoff pulses. If rf and plate pulses are equal in length, phase errors arise at the beginning and end of the pulse due to differences in the rise times of individual amplifier tubes. If, however, the plate pulse is longer than the rf pulse, phase is stable throughout the rf pulse. The switching rate is determined by the duty cycle of the plate pulse and may be made equal to that pulse repetition rate by switching phase in the interpulse period just prior to the transmit pulse.

EVALUATION OF DIFFICULTIES IN OBTAINING STABLE PHASE

The accuracy of the phase-measuring circuits of this system is somewhat dependent on a constant output from the digital phase shifter. Therefore, there must be little or no change in output as various phase combinations are switched. The type GL-6299 tube has a rather wide range of tolerances in its characteristics. A careful selection of the two tubes for a stage was therefore necessary. Differences in power gain of 1 to 2 db per stage have been measured with various tubes in the same strip-transmission-line amplifier.

Although the modified-hybrid power divider shows favorable phase characteristics, it has its failings. The smaller angles of phase shift provide a better match for the preceding stage than do the larger angles. It is, therefore, necessary to tune cathode circuits in conjunction with preceding plate circuits for maximum output and correct phase at the center frequency. This interdependence of successive stages and the large number of resonant circuits (a total of 21) have added to the difficulty of tuning the digital phase shifter.

Once the digital phase shifter is tuned, there is no measurable change in phase with time. The phase shifter of this experiment has been operating well over 1000 hours with no tube failures.

Phase stability is also dependent on the arrangement of the stages, cutoff-pulse magnitude, and drive level. The limits of cutoff-pulse magnitude (6 to 12 volts) and drive (10 to 200 milliwatts peak) have been previously mentioned. Some improvement in the stability of the large-angle phase shifters has been obtained by staggering large and small angles; however, poorer stability occurs in small angles near the end of the amplifier chain. No orderly change in phase has been observed with changes in the above parameters.

The type GL-6299 has a somewhat wide range of tolerances in its characteristics. The approximate value given for C_{gk} is 2.5 to 5.0 $\mu\mu f$ and for C_{gp} , 1.4 to 2.0 $\mu\mu f$ (8). The tuning capacitors, C_T , compensate for variations in these parameters. Because the cathode and plate tap points are tied directly to the lines, differences in the value of C_T and of the interelectrode capacities cause changes in the electrical positions of these tap points. Some control of bandwidth is therefore possible in the plate circuit. In the cathode circuit, however, this phenomenon is likely to cause a mismatch and introduce phase errors. An improved version of the digital phase shifter could be fabricated using a tube such as the G.E. type GL-7644 (9), which is manufactured to closer tolerances than the 6299.

CONCLUSIONS

The L-band digital phase shifter described in this report performs well in a relatively narrow frequency band. It can perform both the transmit and receive phase-shift functions in a narrow-band electronically scanned radar and provide signal amplification in addition.

Within the narrow band limitation, the digital phase shifter is excellent with respect to other parameters. Variations in rf input power over the described range and plate- and cutoff-voltage fluctuations cause negligible phase errors. If plate and cutoff pulses are prepulsed, there is no measurable variation in phase angle throughout the rf pulse. Efforts to provide closely regulated modulators are not necessary. Power supplies and modulators may be of standard design.

ACKNOWLEDGMENT

For origination of the idea and suggestion of development of this device, the authors wish to express grateful appreciation to Mr. I. H. Page, Consultant, Radar Division, NRL.

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